Mixer Linearisation for Software Defined Radio Applications

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***Abstract*—The inherent non-linearity of mixers is particularly acute in broadband receiver design for software defined radio (SDR) applications. Here, the receiver frontend 'sees' not only the wanted channel, but also a number of nearby signals. A conventional mixer will downconvert all of these received channels to IF, thus adding *inband interference* to the wanted channel. In this paper, known mixer linearisation schemes are explained and a new technique using *frequency retranslation* within a linearised mixer architecture is presented. Two-tone-test results from a prototype offered 33dB reduction in the distortion products and 22dB suppression of adjacent channel interference (ACI) for a π/4-DQPSK modulated carrier. A theoretical analysis is also carried out to demonstrate the amplitude and phase matching requirements of the technique.**

Keywords—mixer linearisation, SDR receiver frontend, feedforward, signal cancellation, frequency retranslation.

I. INTRODUCTION

Mixers are key components in communication systems for frequency translating signals. In receivers, mixers are used for downconverting the received radio frequency (RF) signal to baseband or to an intermediate frequency (IF) for further processing. In transmitters, mixers upconvert the baseband signal to IF or to RF for transmitting via an antenna. However, the inherent non-linearity of mixers in communication systems creates numerous undesired effects like harmonic (HD) and intermodulation distortion (IMD), spreading the spectrum to a wider bandwidth. The HD can be filtered out since it appears at one octave higher frequency than the fundamentals, but this requires *appropriate filtering*, whereas IMD cannot be removed by this means and creates ACI to other nearby channels as well as co-channel interference within the same channel. This is particularly acute for a SDR receiver frontend [1-3] which receives not only the wanted channel, but also a number of nearby signals. A non-linear mixer will downconvert all of these received channels together with the wanted channel to IF. During this frequency translation process *inband interference* will be added to the wanted channel, making it more difficult or even impossible for the receiver to correctly detect the information. This places *demanding filtering* requirements [4, 5] on a broadband receiver frontend to reject the out-of-band unwanted channels (blockers) entering the mixer, hence preventing the generation of sufficient inband power to interfere with the wanted signal. However, strong interfering nearby signals may not be rejected. Also, in a traditional radio application the frequency of transmission and reception will be fixed and the filter parameters will be set only for these known frequencies. This is incompatible with the SDR concept and filtering out the blockers of multiple standards will be difficult, thus a linear mixer is highly desirable.

II. MIXER LINEARISATION SCHEMES

Feedforward has been previously applied to amplifiers [6] yielding significant reduction in IMD products. Applying feedforward to mixers necessitates a different approach, since frequency translation occurs making the generation of the reference and error signals difficult. Considering a receiver, the reference (undistorted clean signal at RF input) and the output signals where the IMD products exist (at IF) are at different frequencies, and thus a direct comparison is not possible. Two feedforward linearisation architectures have been proposed for mixers within radio receiver applications, where the reference signal was frequency translated by a backed-off or a saturated secondary mixer. A review of these architectures is given below. In order to show the actual capabilities of each technique, experimental results and problems associated with their application to a receiver frontend are also included.

A. Feedforward Mixer

In [7] the secondary mixer is backed-off to operate in its linear region as shown in Fig. 1. This mixer downconverts the reference signal to the same IF as the output of the main mixer ideally undistorted, but if such a mixer were available, it would no longer be necessary to linearise mixers. This signal when used as a reference is only an *approximation* to the required reference signal. The output of the main mixer, which includes IMD is coupled and added in anti-phase to the output of the secondary mixer, thus cancelling the fundamental signals. This error signal is also an *approximation* to the required error signal, which is then recombined at the output

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combiner to suppress the IMD at the IF output. According to measured results from a similar prototype at University of Bristol [8], the disadvantage of this architecture is that the signal-to-noise ratio (SNR) of the reference path is significantly reduced since it is operating at a much lower RF power. This adds noise to the main path when the error signal is combined at the output combiner to suppress the IMD, which would make the receiver less sensitive to the received signals and reduce the dynamic range of the receiver. Practical results indicated a 25dB reduction in the third-order IMD (IM3) at the IF output when the prototype was used as a downconverter.

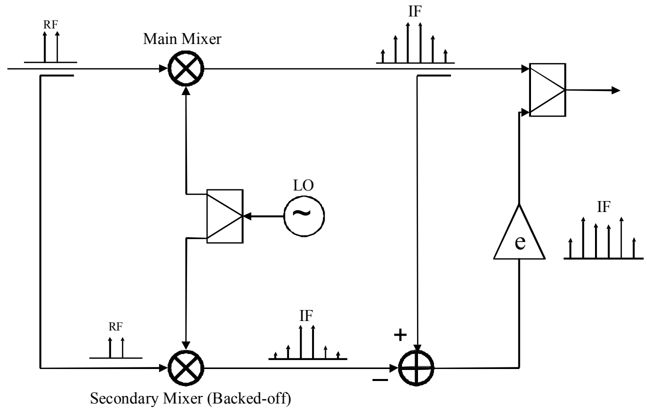


Figure 1: Feedforward error correction.

B. Single-Loop Feedforward Mixer

The addition of noise to the final IF output of the previous configuration (see Fig. 1) was eliminated in the architecture shown in Fig. 2 [9]. Here, the secondary mixer is driven with a much higher RF signal than the main mixer to provide a high level of IMD which is an *approximation* to the required error signal, also providing a high SNR. This error signal is amplitude and phase adjusted before being added to the final IF output for suppressing the IMD. High levels of IM3 reduction can be obtained, about 30dB at a single operating frequency and signal level. This technique offers a low dynamic range, since performance is critically dependent on the amplitude matching of the IMD products and the mismatching characteristics of the two mixers.

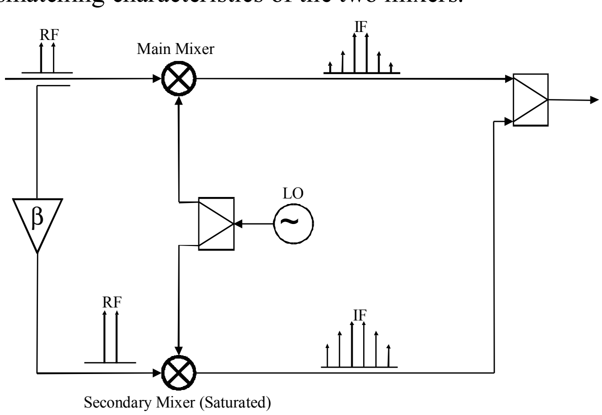


Figure 2: Single-loop feedforward.

III. FREQUENCY RETRANSLATION

Current mixer linearisation techniques are unable to simultaneously offer a large dynamic range, low noise performance and suppress IMD. Fig. 3 shows a novel receiver architecture in order to overcome these shortcomings [10, 11]. The system will be explained considering a receiver application downconverting RF to IF, but it can also be applied to a transmitter.

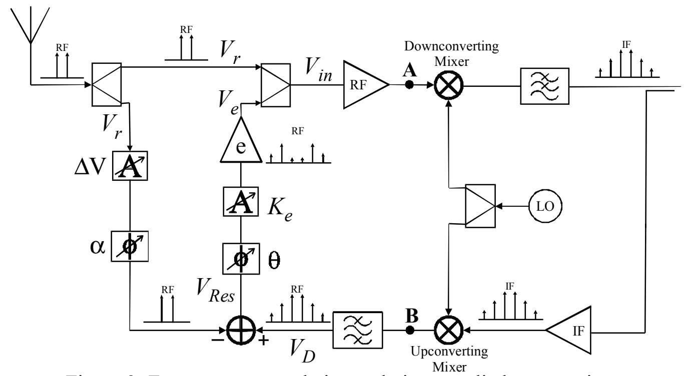
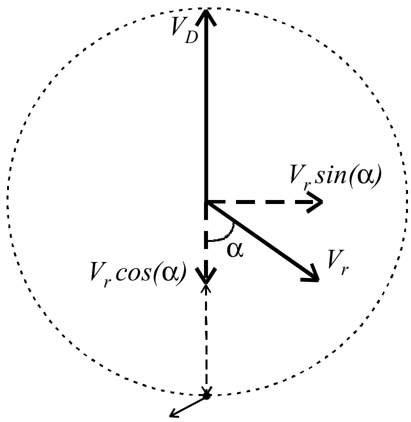


Figure 3: Frequency retranslation technique applied as a receiver.

The distorted output of the downconverting mixer at IF is coupled. This IF output is amplified, *frequency retranslated* back to RF by the upconverting mixer and filtered to remove unwanted image signals. The clean (reference) signal at the receiver frontend is also coupled and added in anti-phase to the frequency retranslated sample of the IF output (which is now at RF) with amplitude correction. This process cancels the fundamental signals and produces an error signal including only the IMD products. This error signal is then combined with the received RF input signal with correct amplitude and phase relation to *predistort* the saturated downconverting mixer. This provides suppression of the IMD without affecting the fundamental signal level, if the signal cancellation is also correctly optimised. The linearity of the second (upconverting) mixer is not so critical since it is not frequency translating the reference signal, but the already distorted IF output. Here, signal cancellation is the vector addition of the reference and frequency retranslated IF output, with system performance critical on the optimisation of this parameter, in common with other feedforward linearisation architectures. Also, after achieving reference signal cancellation, suppressing the IMD products requires predistorting the downconverting mixer with an error signal. These mechanisms will be described in more detail below.

A. Signal Cancellation

The first function within frequency retranslation system is the suppression of the reference signal vector (V_r) at RF by the anti-phase and equal magnitude combination of frequency retranslated IF output (V_D). This is a vector addition and it can be mathematically defined. In Fig. 4, two phasors of unequal amplitude and arbitrary offset from a perfect anti-phase are illustrated.



Perfect magnitude and phase match

Figure 4: Vector addition of reference and frequency retranslated IF output.

Here, the phase adjustment is carried out on the vector V_r, which can be resolved into its orthogonal components. One of these components is anti-phase and the other is orthogonal to V_D. The anti-phase component, $V_r \cos(\alpha)$ will add to V_D destructively and achieve the cancellation of reference signals as shown in (1), where the V_{D2} is the resulting vector.

$$V_{D2} = V_D - V_r \cos(\alpha) \tag{1}$$

The remaining orthogonal vectors V_{D2} and $V_r \sin(\alpha)$ can be expressed using Pythagoras theorem, yielding the final resultant vector V_{Res} as:

$$\begin{aligned} V_{Res}^2 &= V_{D2}^2 + (V_r \sin(\alpha))^2 \\ &= (V_D - V_r \cos(\alpha))^2 + (V_r \sin(\alpha))^2 \end{aligned} \tag{2}$$

$$V_{Res} = \sqrt{(V_D - V_r \cos(\alpha))^2 + (V_r \sin(\alpha))^2}$$

From Fig. 3 it can be seen that the amplitude adjustment is also carried on the reference signal, therefore it can be defined in terms of V_D with a voltage offset of ΔV, as $V_r = V_D + \Delta V$ giving:

$$V_{Res} = \sqrt{(V_D - (V_D + \Delta V)\cos(\alpha))^2 + ((V_D + \Delta V)\sin(\alpha))^2} \tag{3}$$

Normalising to V_D yields:

$$V_{Res} = \sqrt{\left(1 - \left(1 + \frac{\Delta V}{V_D}\right)\cos(\alpha)\right)^2 + \left(\left(1 + \frac{\Delta V}{V_D}\right)\sin(\alpha)\right)^2} \tag{4}$$

and the suppression (V_{Sup}) referred to V_D is:

$$V_{Sup} = 1 - V_{Res} \tag{5}$$

A suppression of unity represents there is no output reference signal magnitude. Equation 5 can be used to calculate the signal suppression ratio for varying magnitude and phase offsets. Fig. 5 shows the magnitude and phase matching requirements for different values of suppression from 50dB to 25dB in increments of 5dB. As the suppression increases, the performance is highly dependent on the magnitude/phase accuracy. However, at lower degrees of signal suppression, variation in magnitude/phase mismatch results in smaller degradation in the signal cancellation performance indicated by the widening of the traces. In order to achieve a signal suppression of 45dB, magnitude and phase match should be better than 0.05dB and 0.5° respectively. These matching requirements are similar to a signal cancellation loop within a feedforward amplifier [12, 13].

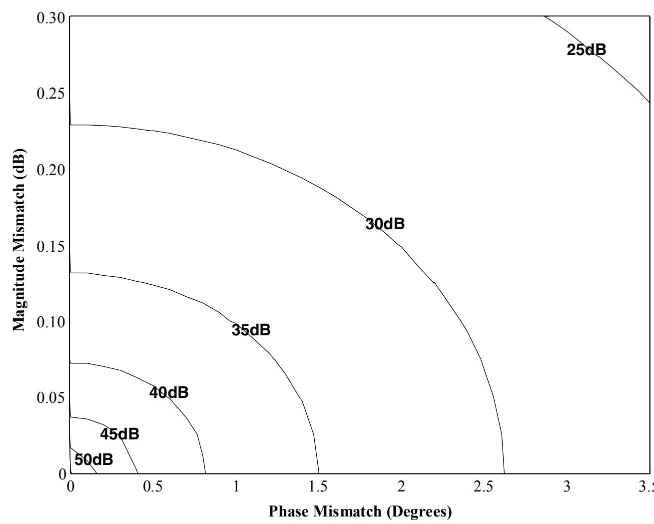


Figure 5: Magnitude and phase match requirement for various levels of signal suppression.

1)Predistortion

After the signal cancellation of the fundamental signals, suppressing the IMD products will be achieved by predistorting the downconverting mixer with the error signal. Assuming that signal cancellation was successful, the gain and phase accuracy of the error signal will determine the IM3 suppression performance, which is similar to the matching requirements of the signal cancellation loop, thus it is not repeated. Here, the system performance is critical on the optimisation of the both signal cancellation and error signal predistortion loop. Therefore, in common with other feedforward circuits, this system is equally sensitive to imperfections. For maintaining the IMD suppression in a practical system, an adaptive control scheme is necessary in order to maintain system performance with changing circuit parameters and input signal conditions.

B. Implementation and Practical Results

A prototype demonstration system is shown in Fig. 6, where two passive double-balanced SRA-2000 Mini-Circuits mixers were used [14]. The RF amplifier preceding the downconverting mixer is offerring a high 1dB gain-compression point, thus driving the downconverting mixer to saturation without adding any additional distortion itself. This is to ensure that the technique is correcting the non-linearity of the mixer and not other circuit elements. The error and IF amplifiers provide sufficient gain at their operating frequencies to compensate for the losses such as coupling, power splitting/combining, filtering and the conversion losses of the mixers. Further, the error amplifier is also operating in its linear range, thus not distorting the error signal. The amplitude and phase adjustment of signal cancellation and predistortion loops has been performed using voltage variable components.

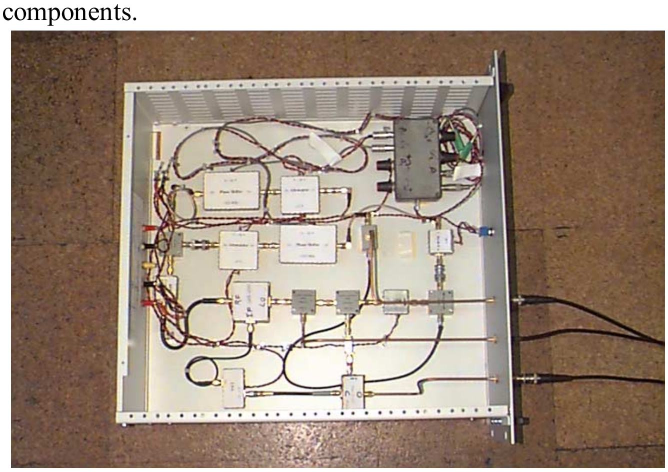


Figure 6: Plan view of the frequency retranslation prototype.

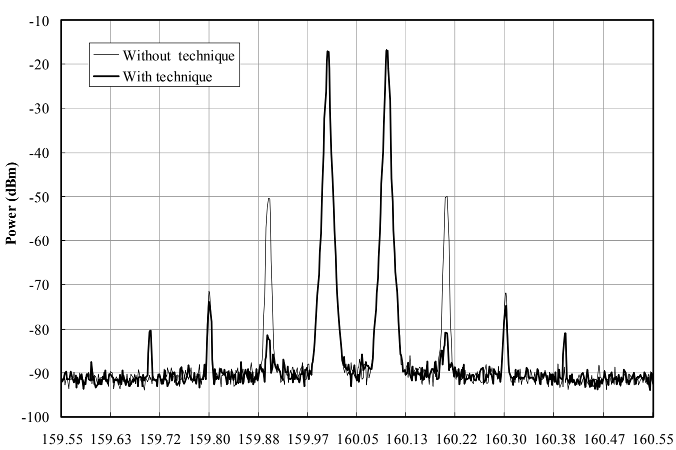


Figure 7: Measured two-tone-test showing a maximum 33dB IM3 improvement with Δf=100kHz.

A two-tone-test was applied at 920MHz with a tone separation (Δf) of 100kHz to provide a downconverted signal at an IF of 160MHz. The IF output of the downconverter with and without the technique applied is given in Fig. 7, indicating an impressive 33dB suppression of IM3. Further amplitude and phase adjustment was performed to suppress IM3 to the same level of fifth-order IMD, where the IM3 improvement is 25dB, i.e. the technique has increased the output third-order intercept (TOI) point of the mixer from –0.17dBm to 12.16dBm. The signal cancellation loop provides more than 40dB suppression of fundamentals as shown in Fig. 8. In order to obtain this error signal, the amplitude and phase match should be within 0.1dB and 1° (see Fig. 5). Noise power measurements at the IF output indicate only 0.2dB increase in the noise figure when the linearisation is applied, which is considered to be negligible. This illustrates that the technique does not degrade the noise performance of the receiver and by correct choice of components it can be further minimised. The same prototype was also tested with a TETRA π/4-DQPSK modulated carrier again downconverted from 920MHz to 160MHz, with Fig. 9 showing a 22dB improvement in ACI. Fig. 10 shows the error signal used for obtaining this ACI improvement.

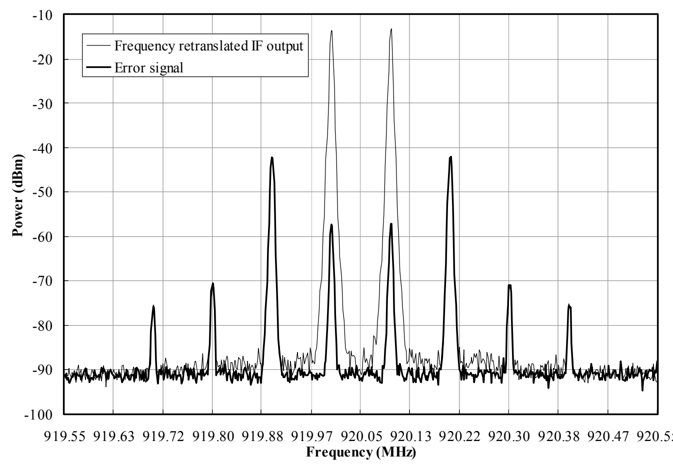


Figure 8: The signal cancellation process with Δf=100kHz.

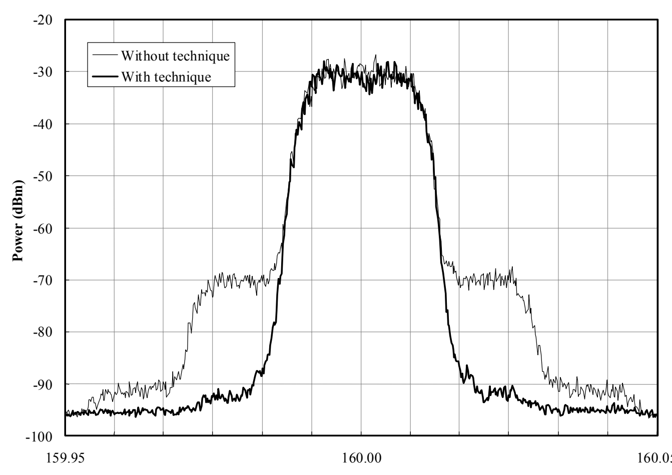


Figure 9: Measured π/4-DQPSK output spectrum showing 22dB ACI suppression.

A two-tone-test was also applied at 920MHz with Δf=500kHz and downconverted to an IF of 160MHz. After applying the proposed technique, 19dB IM3 suppression was obtained. Increasing the frequency separation degrades the signal cancellation and hence the IM3 suppression. Due to the delay mismatch between the two paths in signal cancellation loop, it is not possible to maintain the required 180° phase difference for ideal cancellation at all frequencies. However, it is possible to match the phase at one frequency (in our prototype this is 920MHz), where the perfect cancellation will occur. As the signal frequency deviates from the centre frequency, the cancellation will degrade and reoptimisation will be required to maintain the perfect cancellation. This relationship is measured and shown in Fig. 11. At 920MHz the signal cancellation is about 88dB (Marker 1) and at 2.5MHz offset it reduces to 27.8dB (Marker 2).

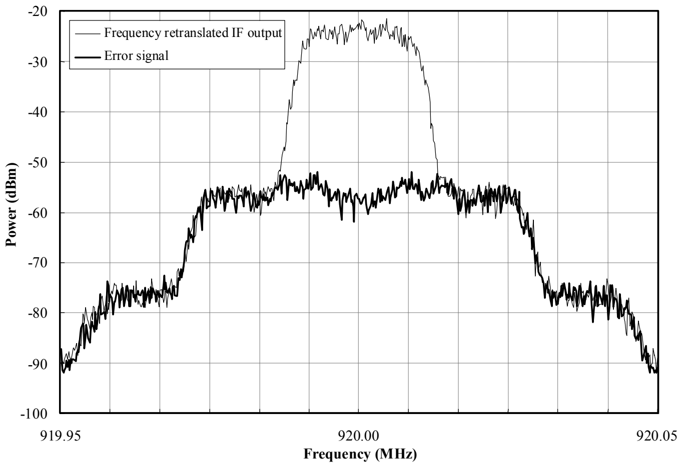


Figure 10: The signal cancellation process with π/4-DQPSK modulated carrier.

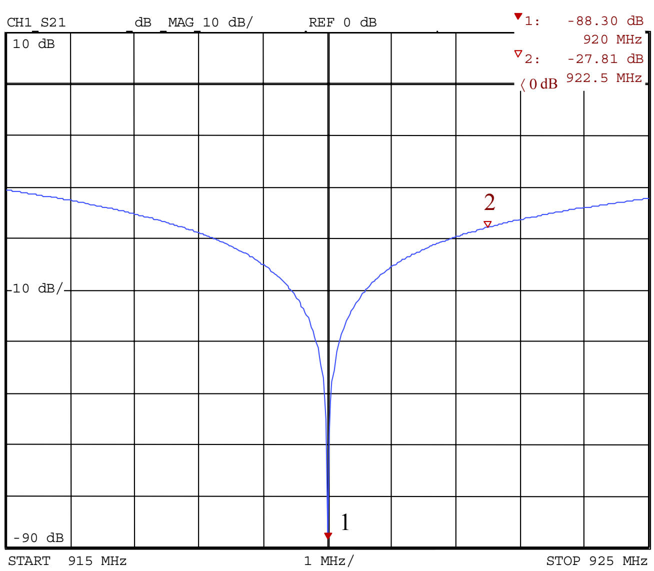


Figure 11: The signal cancellation degrading due to the delay mismatch as the signal frequency deviates from the centre frequency.

IV. Conclusion

In this paper, previously investigated mixer linearisation techniques are summarised and a new technique is presented. A hardware prototype was constructed and the technique was evaluated by practical means. It provides considerable improvement of mixer non-linearity without compromising the noise performance. Measurements indicate suppression of IM3 by up to 33dB, with average suppression of 25dB can be obtained. At this operating point, the calculations show that the output TOI point of the mixer has been increased from -0.17dBm to 12.16dBm. The tests with π/4-DQPSK modulated carrier has shown 22dB ACI improvement at the IF output. Theoretical analysis demonstrates that for a high level IMD suppression, accurate amplitude/phase match is required. The future work will focus on improving the linearisation bandwidth and dynamic range as well as an adaptive control scheme for a practical application of this technique.

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